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2. Patent application number *(The Patent Office will fill in this part)*	9907354.6	30 MAR 1999
3. Full name, address and postcode of the or of each applicant *(underline all surnames)*	University of Bristol Senate House Tyndall Avenue Bristol BS8 1TR UK	
Patents ADP number *(if you know it)* If the applicant is a corporate body, give the country/state of its incorporation	United Kingdom 7981001	
4. Title of the invention Adaptive Filter Equalisation Techniques		
5. Full name of your agent *(if you have one)*	Haseltine Lake & Co.	
"Address for service" in the United Kingdom to which all correspondence should be sent *(including the postcode)*	Imperial House 15-19 Kingsway London WC2B 6UD	
Patents ADP number *(if you know it)*	34001	

	Country	Priority application number *(if you know it)*	Date of filing *(day/month/year)*
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ADAPTIVE FILTER EQUALISATION TECHNIQUES

The present invention relates to adaptive filter equalisation techniques, and in particular to initialisation of coefficients of equalisation filters.

BACKGROUND OF THE PRESENT INVENTION

Modern communication systems utilising wideband communication paths can suffer from a multipath phenomenon known as ISI (inter symbol interference) which reduces the received signal quality. For example in a mobile communication system, schematically shown in Figure 1, the base station 1 transmits an RF signal to a mobile station 2. If the communication is over a direct path, 3, there is no ISI. However, in practical systems reflections occur, from buildings etc. (illustrated by the reflective paths 4 in the figure), causing multiple signals to be received by the receiver. Since direct and reflected path lengths are different, a signal representing a single transmitted symbol (a single bit of data) can arrive at different times. This results in a spreading of the signal for that symbol, referred to as inter-symbol-interference (ISI), which potentially causes difficulty in detecting the symbol.

Figure 2 illustrates an equivalent effect in a wired or fixed communication system where two nodes 5 and 6 are linked. If an intermediate node 7 exists (which connects the path to another node 8) multiple paths 10 and 11 can result. An example of this is communication through cable networks that involve the mains local area network (M-LAN), which can be used in a home-based application. The mains power cables are used to implement a local network in a frequency-isolated area, for example a home or office department. The observed multipath propagation is similar to that shown figure 2, but the interference level could be

much higher for high-speed data communications. This type of communication system is new, but has a very wide application area.

Multipath propagation or time dispersion can also occur, as illustrated in figure 3, in an optical cable 12 with reflection paths 14 causing inter-symbol-interference on the centre trace beam 13.

The multipath channel can be modelled by a tapped delay line filter, as shown in figure 4, where T_s is the symbol period. The resulting signal v_k is made up of the combination of the input symbols x_k, x_{k-1}, $x_{k-2}, \cdots$ x_{k-L} multiplied by filter coefficients h_0, $h_1, \cdots$ h_L and a noise component η_k, as shown in Eq. 1 where $L+1$ is the number of taps in the channel model (or the symbol storage capacity of the channel).

$$v_k = \sum_{i=0}^{L} h_i x_{k-i} + \eta_k \qquad (1)$$

The power delay profile of the channel is shown in figure 5. It can be seen that the channel signal can be modelled as a series of time spaced samples. Since the received signal is spread over a number of symbol periods, then symbols can interfere with one another, giving the phenomenon known as inter-symbol-interference (ISI). This results in the received signals badly defining the original transmitted symbol stream. In the past, various methods have been employed to remove this ISI.

A channel matched filter, such as that shown in figure 6, can be used to convert the channel signals into a waveform, such as that shown in figure 7. The channel-matched filter CMF receives the channel waveform from the channel model, and

combines the received signals with the complex conjugates of the tapped delay line filter model coefficients. The output of the CMF is calculated as,

$$y_k = \sum_{i=0}^{L} h_{L-i}^{*} v_{k-i} \qquad (2)$$

The output ISI profile of the CMF is calculated according to the following equation:

$$d_j = \sum_{i=0}^{L+j} h_i h_{i-j}^{*} \qquad j = -L, \cdots, -1 \qquad (3)$$

$$d_j = \sum_{i=0}^{L} h_i h_i^{*} \qquad j = 0 \qquad (4)$$

$$d_j = \sum_{i=0}^{L-j} h_i h_{i+j}^{*} \qquad j = 1, \cdots, L \qquad (5)$$

Equations (3), (4) and (5) can be expanded for a 5 tap channel profile ($L+1=5$) as,

$$
\begin{aligned}
d_{-4} &= h_0 h_4^* \\
d_{-3} &= h_1 h_4^* + h_0 h_3^* \\
d_{-2} &= h_2 h_4^* + h_1 h_3^* + h_0 h_2^* \\
d_{-1} &= h_3 h_4^* + h_2 h_3^* + h_1 h_2^* + h_0 h_1^* \\
d_0 &= h_4 h_4^* + h_3 h_3^* + h_2 h_2^* + h_1 h_1^* + h_0 h_0^* \\
d_1 &= h_4 h_3^* + h_3 h_2^* + h_2 h_1^* + h_1 h_0^* \\
d_2 &= h_4 h_2^* + h_3 h_1^* + h_2 h_0^* \\
d_3 &= h_4 h_1^* + h_3 h_0^* \\
d_4 &= h_4 h_0^*
\end{aligned}
\qquad (6)
$$

The profile (Figure 7)from the CMF has the advantage of being symmetrical and exhibiting a large real centre tap. It has been shown in several publications that the CMF filter provides the optimum symbol-synchronisation point and multipath diversity at the centre tap, d_0. The output of the CMF can then be fed as an input to a decision feedback equaliser (DFE) filter to remove side lobes calculated by equations (3) and (5). For this type of equalisation a novel direct coefficient calculation method has been developed in the University of Bristol and presented in the MoMuC-2 and VTC'98 conferences. The method, which is called CMF-DFE, calculates the equaliser coefficients directly from the channel profile and produces the best performance for an equaliser filter. However, the required extra unit for running the CMF filter, the feedback data gain adjustment problem and, more importantly the need for a DSP to execute the required Teoplitz Matrix inversion, make the method too expensive and complex for most low-cost, low-power applications.

Alternatively, a minimum mean square decision feedback equaliser (MMSE-DFE) can be used. The MMSE-DFE does not require the CMF filter but suffers from a feedback data gain adjustment problem. The method has a higher bit-error-rate and greater complexity than the CMF-DFE. In these studies the training method is

not adaptive and the training is not suitable for a hardware filter implementation due to the difficulty of matrix inversion.

The least mean square (LMS) algorithm is preferred for most applications that require simple implementation and is also more stable than any other technique. The LMS algorithm suffers when the channel has a null point close to the centre of the unit circle in the z-domain. This results in a slow convergence speed and thus a long training sequence is required. In the standard LMS algorithm, if the training step size is large then the LMS training becomes unstable. If the step size is small then the algorithm may not converge completely by the end of the training sequence.

Although the initialisation technique presented here is explained assuming the LMS training algorithm, the initialisation process is valid for all training algorithms. The accuracy of the initialisation may well allow the feedback filter (FBF) coefficients to be left out of the training process for some applications, with training restricted to the feedforward filter (FFF) coefficients.

DECISION FEEDBACK EQUALISER (DFE)

Figure 8 shows a decision feedback equaliser incorporating a feedforward filter (FFF) (16) and a feedback filter (FBF) (17). $L_{ff}+1$ is the number of feedforward coefficients (18) and L_{fb} is the number of feedback filter coefficients (19). The variables $\hat{x}_k$ (24) and $\bar{x}_k$ (25) represent the estimated data of the DFE and the detected data from the estimated data respectively The feedforward filter (16) comprises a delay line (15), from which signals are tapped, the tapped signals represent the figure 4 samples. The feedback filter (17) also comprises a delay line (20), which has as its input the output of the equaliser $\bar{x}_k$ (25) or the reference training symbols x_k (21). This output is the estimate from the filter of the symbol

concerned. The feedback filter also has a number of taps, which represent the precursor samples.

The tapped signals from both the feedforward and feedback filters are scaled by respective coefficients $c_{-L_{ff}}$ to c_0 (18) and c_1 to $c_{L_{fb}}$ (19), and are then added together to provide the output (24). The expression for the DFE operation is given by

$$\hat{x}_k = \sum_{i=-L_{ff}}^{0} c_i v_{k-i} + \sum_{i=1}^{L_{bb}} c_i x_{k-i} \qquad (7)$$

The output of the equaliser is then detected to determine its level (25). During a training period, usually at the start of the data transmission, the coefficients of the filter can be adjusted so that the output of the equaliser is the expected value. The error, ε_k is calculated for use in the training algorithm as

$$\varepsilon_k = x_k - \hat{x}_k \qquad (8)$$

The coefficients can be obtained in various manners using so-called adaptive training techniques. These adaptive training techniques use a known training sequence of symbols, which allow the receiver to compare the detected symbol sequence with the expected sequence, as shown in Eq.(8). The equaliser coefficients can then be adjusted until the detected symbols have an error rate within the required tolerance (or until the end of the training sequence is reached).

As mentioned above, the coefficients can be calculated in several ways. The various methods can be grouped into linear (such as least mean squares) and non-linear (such as recursive least squares RLS) techniques. The least mean squares (LMS) algorithm is the simplest method for equaliser training. However, the

performance of conventional LMS algorithms is quite poor and its convergence rate is slow. The recursive least squares algorithm provides very high performance but is complex to implement. It is also possible to directly calculate the equaliser coefficients. Examples of direct calculation methods include the minimum mean square error and channel matched filter equaliser.

The main limitations of direct coefficient calculation are the need to estimate the required channel accurately and the problem of estimating the required feedback data gain in order to adapt the incoming data level to the detector output level. Both of these problems require powerful high-speed microprocessors to execute the required matrix inversion within realistic timescales.

SUMMARY OF THE PRESENT INVENTION

According to one aspect of the present invention, there is provided a method of estimating coefficients of an equaliser filter which receives an input channel signal, the method including initialising preselected coefficients of the equaliser filter in accordance with a calculated channel matched filter model, and estimating the coefficients of the filter using an estimation technique and a known input channel signal to the equaliser..

BRIEF DESCRIPTION OF THE DRAWINGS

Figures 1, 2 and 3 illustrate multi-path signal transmission in communication systems;

Figure 4 is a block diagram illustrating a model of the multi-path signal transmission channel of figures 1, 2 and 3;

Figure 5 illustrates the power delay profile of the channel and sampling points for the channel coefficients modelled in the figure 4;

Figure 6 illustrates a channel matched filter (CMF);

Figure 7 illustrates the output power delay profile or ISI profile of the CMF;

Figure 8 illustrates the decision feedback equaliser (DFE);

Figures 9 and 10 illustrate relative performance characteristics of known equalisation techniques and an equalisation technique incorporating a method embodying the present invention.

DESCRIPTION OF THE PREFERRED EMBODIMENT

The present invention is concerned with the calculation of the coefficient values in the feedforward and feedback filters of the decision feedback equaliser such that the approximation of the coefficients of the equaliser can be made more efficient. The method can also be applied to the FFF in a Linear Transversal Equaliser (LTE).

In the particular embodiment to be described, the technique used for actually estimating the coefficients is the least means squares algorithm, however any appropriate estimation algorithm can be used with the initialisation method embodied in the present invention.

As discussed above, the decision feedback equaliser is typically trained using the training data at the start of a communications packet. The training data is known at the receiver allowing the actual detected symbol stream to be compared with the required symbol stream. The equaliser can then adapt to the particular communication path concerned.

In the ordinary case, the strongest path (h_1 illustrated in figure 4) of the channel is chosen for the symbol synchronisation. Pre-cursor path (only h_0 in fig. 4) symbols are cancelled by the feedforward filter (FFF) and post-cursor path (h_2, h_3, h_4)

symbols are cancelled by the feedback filter (FBF). The FFF is an anti-causal, finite impulse response (FIR) filter. For a good solution, a filter size equal to or bigger than the length of the channel model is chosen. The FFF works in order to combine the power delay profile in its targeted window. The FBF is causal and normally implemented using L-1 taps, it cancels the received signal energy for it's targeted window. Therefore, energy in the vicinity of the synchronisation symbol should be reserved in the feedforward filter ISI cancellation window to obtain more multipath diversity.

In order to do this, the least delayed transmit symbol (x_k) at the centre tap data (v_k in figure 8) should be used for symbol synchronisation. All received power for symbol x_k is represented in the feedforward filter and other interfering symbols are shared by the feedforward and feedback filters. Figure 8 is arranged to obtain the symbol synchronisation at symbol x_k, which is received through the first tap of the channel h_0 as it appears in the centre-tap data v_k. Then, subsequent interference symbols x_{k+1} *to* x_{k+L_f+L-1} and previous interference symbols x_{k-1} *to* x_{k-L} are targeted as interference components by the FFF and FBF respectively. Since the FBF is causal, there is no need to employ more than L-1 taps.

When this kind of symbol synchronisation is applied, all adaptive training algorithms (RLS, LMS etc.) approximate the equaliser coefficients slowly and require many training iterations. If the convergence speed is increased by either incrementing the training step size of the LMS algorithm or by decrementing the forgetting factor of the RLS algorithm then the algorithms become unstable with this type of symbol synchronisation. When symbol synchronisation is performed on the strongest path of the channel, ISI cancellation is easy and training becomes stable hence the ISI components have lower energy than the targeted symbols.

However, their approximation does not maximise the multipath diversity profit of the channel.

When the channel power delay profile is estimated during the frame synchronisation process the channel coefficients are used to implement a CMF filter in the FFF of the DFE, which are expressed in equation (9).

$$c_j = h_{-j}^{*} \text{ where } j = -L, \cdots 0$$

$$c_j = 0 \text{ where } -L_{ff} \leq j < -L \text{ (assuming } L_{ff} \geq L\text{)}$$

(9)

Then the FFF filter coefficients for a 6 tap FFF and 5 tap channel model appear as

$$c_{-5} = 0$$
$$c_{-4} = h_4^*$$
$$c_{-3} = h_3^*$$
$$c_{-2} = h_2^*$$
$$c_{-1} = h_1^*$$
$$c_0 = h_0^* \quad \text{(centre tap)}$$

(10)

These coefficients implement the CMF filter in the FFF. As mentioned above, the CMF extends the interference profile as shown in Figure 7 and presents all the multipath energy for the desired symbol at the centre ISI component d_0. The above initialisation secures all the received signal energy about the target symbol x_k in the feedforward filter. This is a unique starting point for a linear transversal equaliser (LTE) and is not dependent on any particular adaptive training algorithm.

When the operation of the DFE is considered, the feedback coefficients can be initialised according to the ISI profile of the CMF. The FBF works as a substructure of the target interference component and the previous symbols ISI components should be subtracted from the estimate of the DFE by the FBF. According to this phenomena, the FBF coefficients would be given as shown in equation (11)

$$c_j = -d_j \text{ where } j = 1, \cdots L \tag{11}$$

and the FBF coefficients for a 5-tap channel profile would be as

$$\begin{aligned} c_1 &= -d_1 \\ c_2 &= -d_2 \\ c_3 &= -d_3 \\ c_4 &= -d_4 \end{aligned} \qquad (12)$$

Although the above initialisation is valid for a unit-valued received signal level, most of the time the received signal has non-unity level and is not compatible with the unit-valued detected data in the FBF. Therefore the initialisation coefficients should be normalised in order to provide compatibility. It is also presented here that the centre tap of the CMFs' ISI profile d_0 can be used for this propose as shown in equation (13).

$$c_{j,NORMALISED} = \frac{c_j}{d_0} \text{ where } j = -L, \ldots\ldots -1, 0, 1, \ldots\ldots L \qquad (13)$$

Therefore if a (6,4) DFE, 6 tap FFF and 4 tap FBF is the concern for equalising a 5 tap channel then the DFE should be initialised as

$$
\begin{aligned}
c_{-5} &= 0 \\
c_{-4} &= h_4^* / d_0 \\
c_{-3} &= h_3^* / d_0 \\
c_{-2} &= h_2^* / d_0 \\
c_{-1} &= h_1^* / d_0 \\
c_0 &= h_0^* / d_0 \\
c_1 &= -d_1 / d_0 \\
c_2 &= -d_2 / d_0 \\
c_3 &= -d_3 / d_0 \\
c_4 &= -d_4 / d_0
\end{aligned}
\qquad (14)
$$

When the initialisation is in accordance with equation (14), the training would start with a proper ISI cancellation on the previous symbols side of the cursor symbol and the remaining ISI components would be on the left side (subsequent symbols) of the CMFs' power delay profile as shown in Figure 7. In this case the FBF coefficients can keep these values and the FFF coefficients are forced to implement the ISI cancellation according to this pre-determined FBF setting. However, including the FBF in the training process does increase performance since updated FFF filter coefficients require a different set of FBF filter coefficients. Simulations have shown that the FBF coefficients do not significantly change, because the error function does not have a large effect on the FBF since the FBF operation is initially correct.

This feature of the FBF, and the initialised values of the FFF coefficients, holds the training algorithm in a stable mode allowing the training algorithm to use a

bigger step size or smaller forgetting factor for calculating the FFF coefficients. An adaptive training algorithm with a short training sequence is more desirable than one with a long training sequence and non-adaptive equalisation. The presented invention reduces the required training sequence length dramatically, since the initialised coefficients are close to the final values achieved after the iterative training.

The features of this initialisation are now studied with an LMS algorithm, which is a member of the stochastic gradient-based algorithms, acting as an example. It uses an estimate of the gradient of the error function and as such does not require a measurement of the pertinent correlation functions (nor does it require matrix inversion). The LMS algorithm is simple and robust and performs well under a wide range of channel conditions and input signal powers.

In order to determine the values of the tap coefficients required for a given channel, a training sequence of data bits is supplied to the receiver. In a typical LMS system this can be anything up to 800 bits of information. Since the receiver knows the training sequence, the output from the DFE can be compared with the expected output and the coefficients adjusted accordingly. The update equation for the ordinary LMS algorithm is given as

$$\begin{aligned} c_{i,k+1} &= c_{i,k} + \Delta.\varepsilon_k.v_{k-i}^{*} \text{ when } i = -L_{ff},\cdots\ 0 \\ c_{i,k+1} &= c_{i,k} + \Delta.\varepsilon_k.x_{k-i}^{*} \text{ when } i = 1,\cdots\ L \end{aligned} \qquad (15)$$

where k is the time index, Δ is the step-size of the LMS algorithm and ε_k the error as calculated at the k-th internal (equation (8)). $(^{*})$ represents the complex conjugate of the variable. It has been suggested in several textbooks that the step-size for the LMS in equalisation training should be equal to between 0.005 and 0.1, with 0.045 quoted for stable training. The LMS algorithm converges within

400-800 training iterations. When the presented initialisation is applied to the LMS algorithm the update equation becomes

$$\begin{aligned} c_{i,k+1} &= c_{i,k} + \Delta_{ff} . \varepsilon_k . v^*_{k-i} \text{ when } i = -L_{ff}, \cdots 0 \\ c_{i,k+1} &= c_{i,k} + \Delta_{fb} . \varepsilon_k . x^*_{k-i} \text{ when } i = 1, \cdots L \end{aligned} \qquad (16)$$

where the step sizes for the FFF and FBF respectively Δ_{ff} and Δ_{fb} differ from the original update equation. The same DFE filter is used. Simulation studies have shown that the FFF step size Δ_{ff} can be increased to 0.2 without causing any instability. The feedback filter step size Δ_{fb} should be within the range 0 to 0.1, a typical value is 0.01. When the FBF step size Δ_{fb} is equal to 0 it means the FBF filter is removed from the training process. This does not cause any major performance degradation, in the Supervised LMS, and the error performance is still better than the ordinary LMS algorithm, as shown in the performance curves of figures 9 and 10. The bigger step size for the FFF, of 0.2, dramatically reduces the required training iterations to 100-150 for a reasonable low mean square error value as shown in Figure 9.

Figures 9 and 10 are graphs illustrating the relative performance of the training algorithms. It can be seen that the initialisation technique embodied in the present invention greatly improves the speed of convergence as well as significantly reducing the bit-error-rate. It is clear in figure 9 that the equaliser starts with a lower mean square error value and reduces this to a lower rate within 150 iterations. These curves are produced for fixed step size training. If variable step size techniques were applied, as mentioned in several patent applications, the mean square error value could be improved resulting in a lower bit-error-rate (BER). In figure 9 the FBF step size is kept constant at Δ_{fb} =0.02 and the FFF step

size Δ_{ff} is changed. Bigger step sizes for the FFF noticeably reduces the mean square error and does not cause instability.

Figure 10 show the bit-error-rate analysis of the LMS technique with the proposed *Supervised LMS* (SLMS) initialisation. In figures 10.a, 10.b and 10.c both the ordinary and supervised LMS algorithms are trained over 200 and 450 iterations. There is no noticeable benefit from using more training iterations for the SLMS algorithm and in all cases the performance is much higher than the ordinary LMS. When the training for the FBF is ignored there is no important increase in the bit-error-rate. Therefore this method may well be preferred in practice.

If another training algorithm with faster convergence is used, i.e. the RLS algorithm, for the training then the speed improvement would be expected to be similar to the LMS, i.e. three times faster than normal. The performance improvement, in terms of bit-error-rate, should also be observed. However, ignoring the training of the FBF would dramatically reduce the complexity and speed problems of the RLS algorithm thus allowing it to be implemented with reasonable processing power.

As will be readily appreciated, embodiments of the present invention can present a low cost, high speed, high performance technique for equalisation. The only additional processing required is calculating the solution to the equations (4), (5) and (14), a single real valued division, since the centre tap of the CMF d_0 is real valued, and a number of extra multiply-accumulate operations (15 for $L=5$ and 21 for $L=6$). Although these appear as extra system complexity, the equations (4) and (5) can be executed by the same autocorrelation filter used for bit synchronisation and the division operation in equation (14) is a data normalisation process that can be implemented by data shifting operations. Therefore the initialisation method is suitable for either DSP or hardware filter implementations.

CLAIMS:

1. A method of estimating coefficients of an equaliser filter which receives an input channel signal, the method including initialising preselected coefficients of the equaliser filter in accordance with a calculated channel matched filter model, and estimating the coefficients of the filter using an estimation technique and a known input channel signal to the equaliser.

2. A method as claimed in claim 1, wherein the received signal is modelled as a tapped delay line filter model, wherein the channel matched filter input signal is modelled according to the expressions:

$$d_j = \sum_{i=0}^{L+j} h_i h_{i-j}^* \qquad j = -L, \cdots, -1$$

$$d_j = \sum_{i=0}^{L} h_i h_i^* \qquad j = 0 \text{ , and}$$

$$d_j = \sum_{i=0}^{L-j} h_i h_{i+j}^* \qquad j = 1, \cdots, L$$

where d represents the inter symbol interference components of the channel matched filter samples, h the channel filter model coefficients, h* the complex conjugates of h, and L+1 equals the number of taps in the filter model of the channel

and wherein the preselected coefficients are feed forward filter coefficients of the equaliser filter, which coefficients are initialised according to the expression:

$$c_j = h_{-j}^* \text{ where } j = -L, \cdots 0 \text{ ,}$$

c being the coefficient concerned.

3. A method as claimed in claim 1 or 2, wherein the received signal is modelled as a tapped delay line filter model, wherein the channel matched filter input signal is modelled according to the expressions:

$$d_j = \sum_{i=0}^{L+j} h_i h_{i-j}^* \qquad j = -L,\cdots,-1$$

$$d_j = \sum_{i=0}^{L} h_i h_i^* \qquad j = 0\ ,$$

where d represents the inter symbol interference components of the channel matched filter samples, h the channel filter model coefficients, h* the complex conjugates of h, and L+1 equals the number of taps in the filter model of the channel

and wherein the preselected coefficients are feed back filter coefficients of the equaliser filter, which coefficients are initialised according to the expression:

$$c_j = -d_j \text{ where } j = 1,\cdots L$$

where c is the coefficient concerned.

4. A method as claimed in claim 3, wherein the initialised coefficients are normalised with respect to the centre tap channel matched filter sample, according to the following relationship:

$$c_{j,NORMALISED} = \frac{c_j}{d_0} \text{ where } j = -L,\ldots\ldots -1,0,1,\ldots\ldots L \quad .$$

5. A method as claimed in any one of claims 1 to 4, wherein the estimation technique is a least mean square technique.

6. A method as claimed in any one of claims 1 to 4, wherein the estimation technique is a recursive least squares technique.

7. A method as claimed in any one of the preceding claims, wherein the feedforward and feedback filters of a decision feedback filter are trained independently at respective training rate given by the following expression:

$$c_{i,k+1} = c_{i,k} + \Delta_{ff} . \varepsilon_k . v^*_{k-i} \text{ when } \quad i = -L_{ff}, \cdots 0$$
$$c_{i,k+1} = c_{i,k} + \Delta_{fb} . \varepsilon_k . x^*_{k-i} \text{ when } \quad i = 1, \cdots L$$

8. A method as claimed in claim 1, wherein symbol synchronisation is made on the least delayed path of the channel.

ABSTRACT

ADAPTIVE FILTER EQUALISATION TECHNIQUES

Multipath communications channels can result in intersymbol interference that requires the use of equalisation techniques to be removed.

A method of estimating coefficients of an equaliser filter which receives an input channel signal, includes initialising preselected coefficients of the equaliser filter in accordance with a calculated channel matched filter model, and estimating the coefficients of the filter using an estimation technique and a known input channel signal to the equaliser. Initialising the coefficients in this way enables the training period of the filter to be reduced.

[Fig. 1]

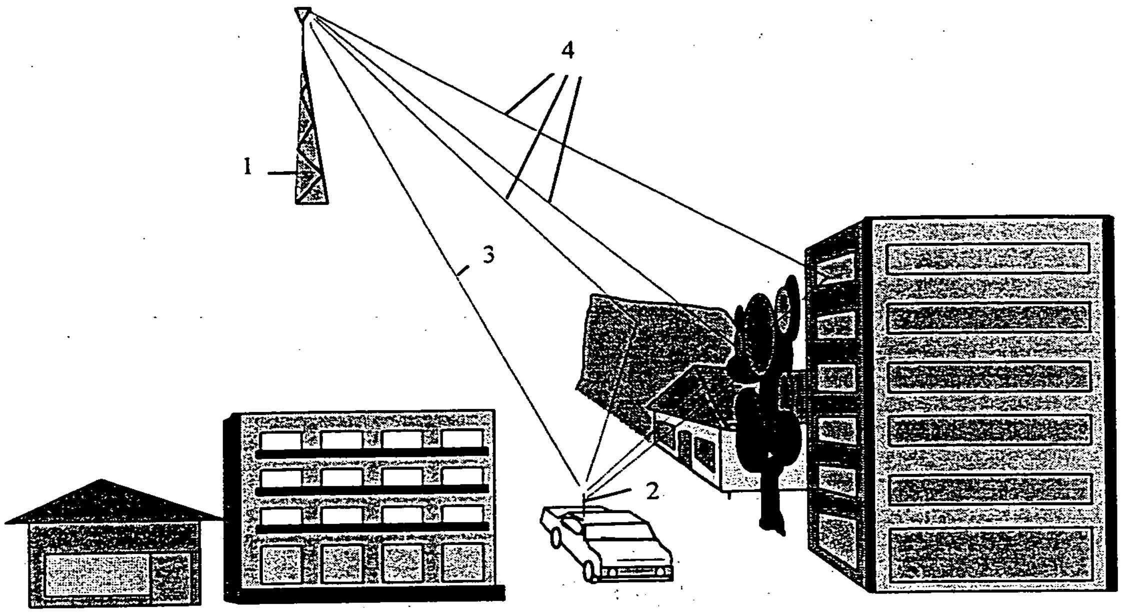

Figure 1.

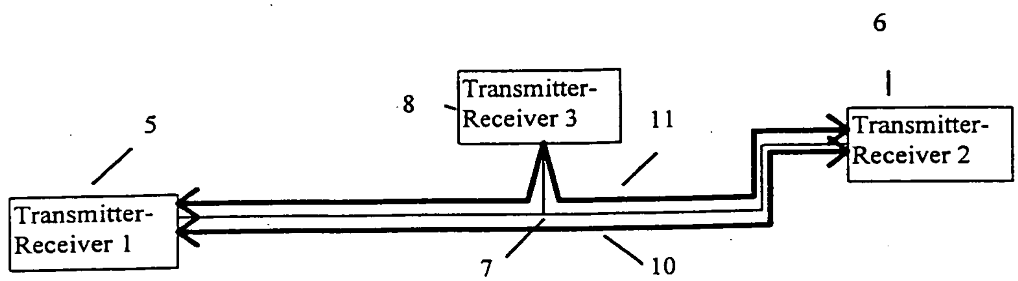
6
8
Transmitter-
Receiver 3
5
11
Transmitter-
Receiver 2
Transmitter-
Receiver 1
7
10

Figure 2.

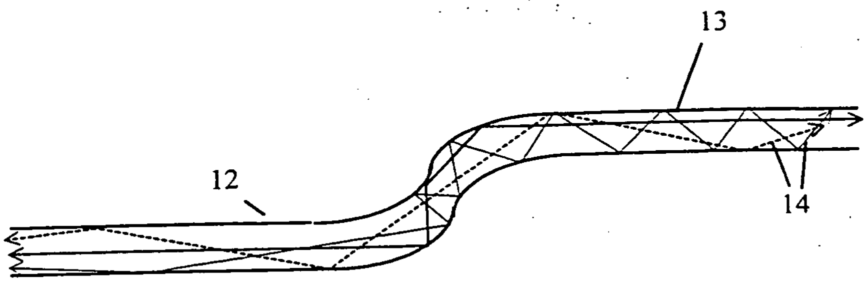

Figure 3.

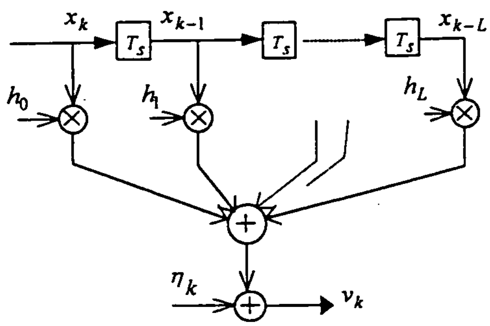

Figure 4.

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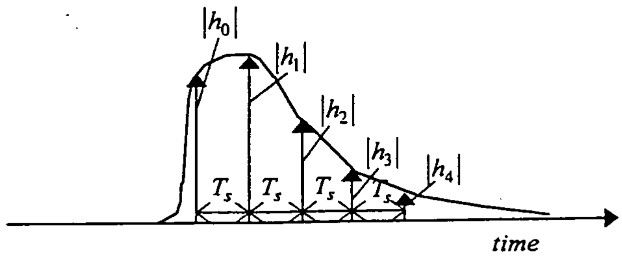


Figure 5.

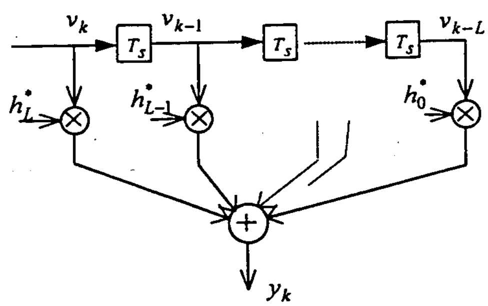


Figure 6.

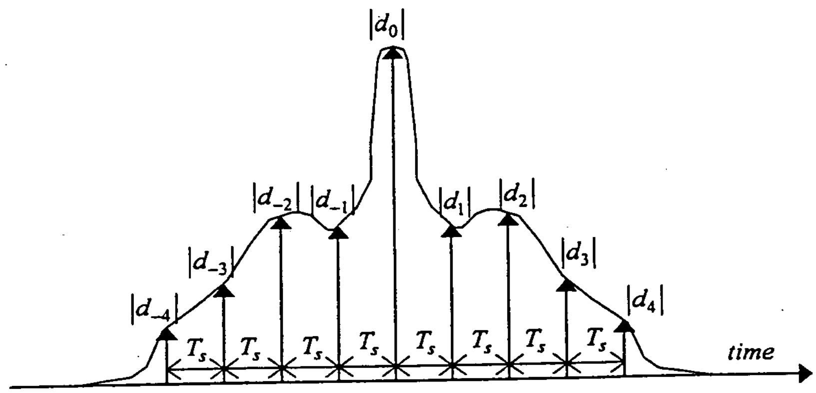
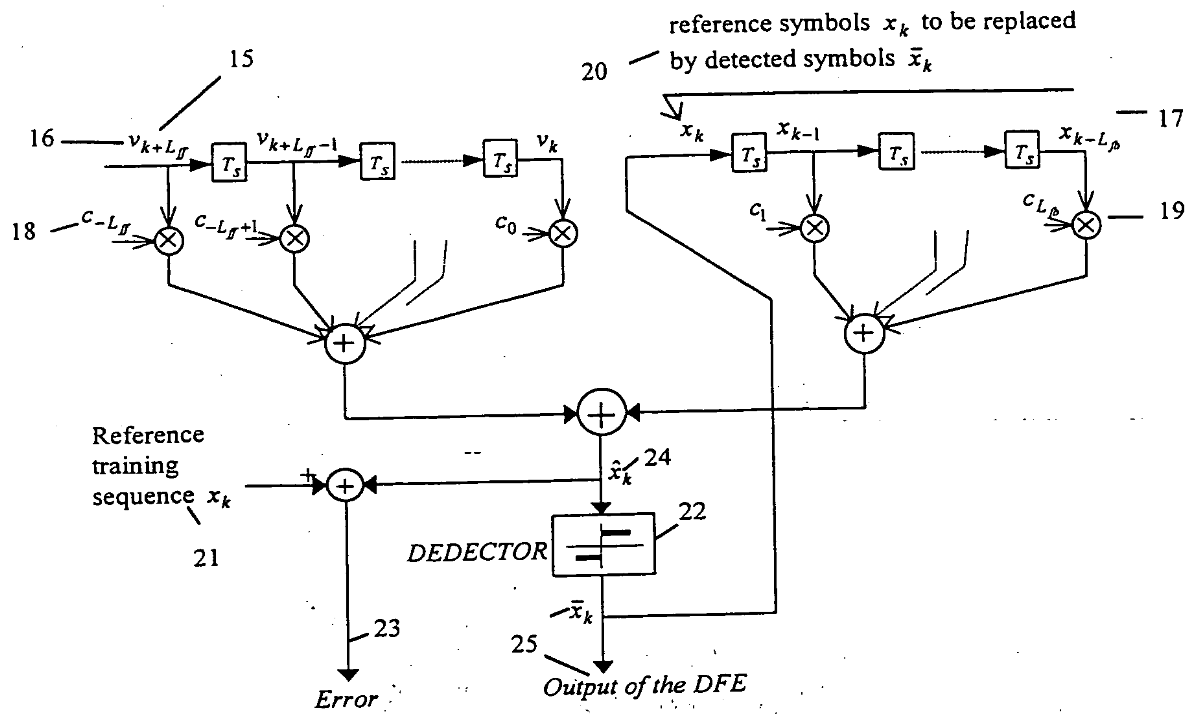


Figure 7.



reference symbols x_k to be replaced by detected symbols $\bar{x}_k$
15
16
20
17
18
19
$v_{k+L_{ff}}$
$v_{k+L_{ff}-1}$
v_k
T_s
$c_{-L_{ff}}$
$c_{-L_{ff}+1}$
c_0
x_k
x_{k-1}
$x_{k-L_{fb}}$
c_1
$c_{L_{fb}}$
Reference training sequence x_k
21
$\hat{x}_k$
24
22
DEDECTOR
$\bar{x}_k$
23
25
Error
Output of the DFE

Figure 8.

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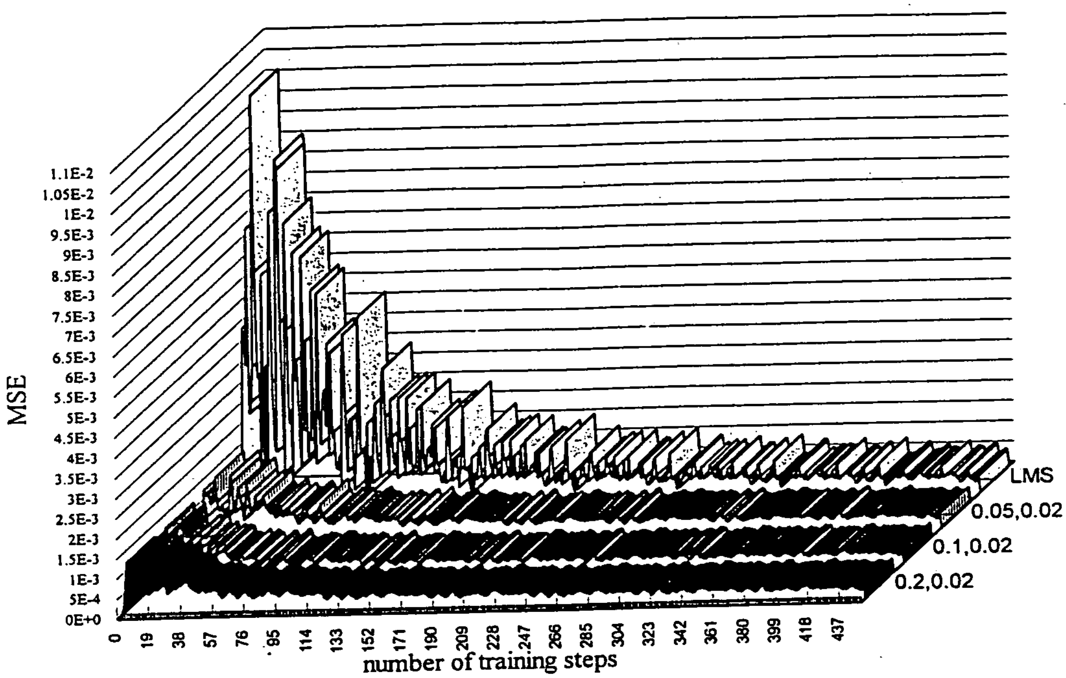

MSE
1.1E-2
1.05E-2
1E-2
9.5E-3
9E-3
8.5E-3
8E-3
7.5E-3
7E-3
6.5E-3
6E-3
5.5E-3
5E-3
4.5E-3
4E-3
3.5E-3
3E-3
2.5E-3
2E-3
1.5E-3
1E-3
5E-4
0E+0
0
19
38
57
76
95
114
133
152
171
190
209
228
247
266
285
304
323
342
361
380
399
418
437
number of training steps
LMS
0.05,0.02
0.1,0.02
0.2,0.02

Figure 9.

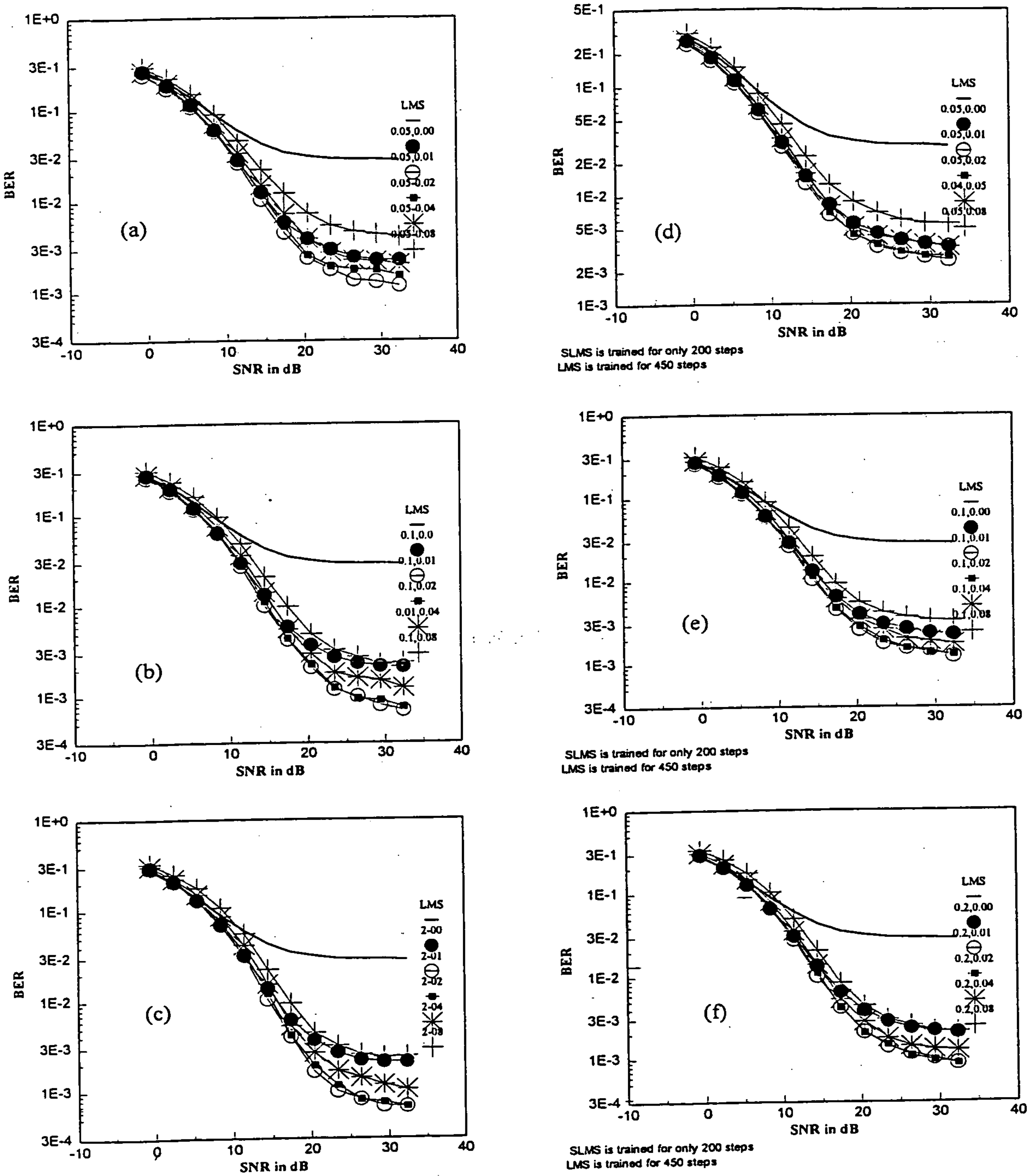
(a)
BER
SNR in dB
LMS
0.05,0.00
0.05,0.01
0.05-0.02
0.05-0.04
0.05-0.08
(b)
BER
SNR in dB
LMS
0.1,0.0
0.1,0.01
0.1,0.02
0.01,0.04
0.1,0.08
(c)
BER
SNR in dB
LMS
2-00
2-01
2-02
2-04
2-08
(d)
BER
SNR in dB
LMS
0.05,0.00
0.05,0.01
0.05,0.02
0.04,0.05
0.05,0.08
SLMS is trained for only 200 steps
LMS is trained for 450 steps
(e)
BER
SNR in dB
LMS
0.1,0.00
0.1,0.01
0.1,0.02
0.1,0.04
0.1,0.08
SLMS is trained for only 200 steps
LMS is trained for 450 steps
(f)
BER
SNR in dB
LMS
0.2,0.00
0.2,0.01
0.2,0.02
0.2,0.04
0.2,0.08
SLMS is trained for only 200 steps
LMS is trained for 450 steps

Figure 10.

PCT APPLICATION NUMBER : PCT/GB00/01214

FILING DATE FORM 23/77 : 30 MARCH 00.

AGENT : HASELTINE LAKE & CO.

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